Polar Codes: Speed of polarization and polynomial gap to capacity

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Abstract—We prove that, for all binary-input symmetric memoryless channels, polar codes enable reliable communication at rates within $\varepsilon > 0$ of the Shannon capacity with a block length, construction complexity, and decoding complexity all bounded by a *polynomial* in $1/\varepsilon$. Polar coding gives the *first known explicit construction* with rigorous proofs of all these properties.

We give an elementary proof of the capacity achieving property of polar codes that does not rely on the martingale convergence theorem. As a result, we are able to explicitly show that polar codes can have block length (and consequently also encoding and decoding complexity) that is bounded by a polynomial in the gap to capacity. The generator matrix of such polar codes can be constructed in polynomial time using merging of channel output symbols to reduce the alphabet size of the channels seen at the decoder.

I. Introduction

In this work, we establish that Arıkan's celebrated polar codes [1] have the desirable property of fast convergence to Shannon capacity. Specifically, we prove that polar codes can operate at rates within $\varepsilon > 0$ of the Shannon capacity of binary-input memoryless output-symmetric (BIS) channels with a block length $N = N(\varepsilon)$ that grows only polynomially in $1/\varepsilon$. Further, a generator matrix of such a code can be deterministically constructed in time polynomial in the block length N. For decoding, Arıkan's successive cancellation decoder has polynomial (in fact $\mathcal{O}(N \log N)$) complexity.

Thus, the delay and construction/decoding complexity of polar codes can *all* be polynomially bounded as a function of the gap to capacity. This provides a complexity-theoretic backing for the statement "polar codes are the first constructive capacity achieving codes," common in the recent coding literature. As explained below, these attributes together distinguish polar codes from the Forney/Justesen style concatenated code constructions for achieving capacity.

Our analysis of polar codes avoids the use of the martingale convergence theorem — this is instrumental in our

Due to space constraints, the full details of some proofs have been omitted from this version. The full version is available on the Electronic Colloquium on Computational Complexity (ECCC) as TR13-050.

polynomial convergence bounds and as a side benefit makes the proof elementary and self-contained.

A. Context

Shannon's noisy channel coding theorem implies that for every memoryless channel W with binary inputs and a finite output alphabet, there is a capacity $I(W) \geqslant 0$ and constants $a_W < \infty$ and $b_W > 0$ such that the following holds: For all $\varepsilon > 0$ and integers $N \geqslant a_W/\varepsilon^2$, there *exists* a binary code $C \subset \{0,1\}^N$ of rate at least $I(W) - \varepsilon$ which enables reliable communication on the channel W with probability of miscommunication at most $2^{-b_W \varepsilon^2 N}$. A proof implying these quantitative bounds is implicit in Wolfowitz's proof of Shannon's theorem [2].

This remarkable theorem showed that a constant factor redundancy was sufficient to achieve arbitrarily small probability of miscommunication, provided we tolerate a "delay" of processing N channel outputs at a time for large enough block length N. Further, together with a converse theorem, it precisely characterized the minimum redundancy factor (namely, $1/I(W)$) needed to achieve such a guarantee. It is also known that a block length of $N \geqslant \Omega(1/\varepsilon^2)$ is required to operate within ε of capacity and even a constant, say 0.1, probability of miscommunication; in fact, a very precise statement that even pinned down the constant in the $\Omega(\cdot)$ notation was obtained by Strassen [3].

As Shannon's theorem is based on random coding and is non-constructive, one of the principal theoretical challenges is to make it constructive. More precisely, the goal is to give an explicit (i.e., constructible in deterministic $\mathrm{poly}(N)$ time) description of the encoding function of the code, and a polynomial time error-correction algorithm for decoding the correct transmitted codeword with high probability (over the noise of the channel). Further, it is important to achieve this with small block length N as that corresponds to the delay at the receiver before the message bits can be recovered.

For simplicity let us for now consider the binary symmetric channel (BSC) with crossover probability p, $0 < p < 1/2$, denoted BSC_p (our results hold for any BIS channel). Recall that BSC_p flips each input bit independently

with probability p, and leaves it unchanged with probability $1 - p$. The Shannon capacity of BSC_p is $1 - h(p)$, where $h(x) = -x \log_2 x - (1-x) \log_2(1-x)$ is the binary entropy function. For the BSC, the capacity can be achieved by binary linear codes.

One simple and classic approach to *construct* capacity-achieving codes is via Forney's concatenated codes [4]. We briefly recall this approach (see, for instance, [5, Sec. 3] for more details). Suppose we desire codes of rate $1 - h(p) - \varepsilon$ for communication on BSC_p. The idea is to take as an outer code any binary linear code $C_{\text{out}} \subset \{0,1\}^{n_0}$ of rate $1 - \varepsilon/2$ that can correct a fraction $\gamma(\varepsilon) > 0$ of worst-case errors. Then, each block of $b = \Theta(\frac{1}{\varepsilon^2} \log(1/\gamma))$ bits of the outer codeword is further encoded by an inner code of rate within $\varepsilon/2$ of Shannon capacity (i.e., rate at least $1 - h(p) - \varepsilon/2$). This inner code is constructed by brute force in time $\exp(\mathcal{O}(b))$. By decoding the inner blocks by finding the nearest codeword in $\exp(\mathcal{O}(b))$ time, and then correcting up to $\gamma(\varepsilon) n_0$ errors at the outer level, one can achieve exponentially small decoding error probability. However the decoding complexity grows like $n_0 \exp(\mathcal{O}(b))$. Thus both the construction and decoding complexity have an exponential dependence on $1/\varepsilon$. In conclusion, this method allows one to obtain codes within ε of capacity with a block length polynomially large in $1/\varepsilon$. However, the construction and decoding complexity grow exponentially in $1/\varepsilon$, which is undesirable.[1]

B. Our result: polynomial convergence to capacity of polar codes

In this work, we prove that Arıkan's remarkable polar codes allow us to approach capacity within a gap $\varepsilon > 0$ with *delay* (block length) and *complexity* both depending polynomially on $1/\varepsilon$. Polar codes are the *first* known construction with this property.[2]

Below is a formal statement of the main result, stated for BIS channels. For general, non-symmetric channels, the same claim holds for achieving the *symmetric capacity*, which is the best rate achievable with the uniform input bit distribution.

Theorem 1. *There is an absolute constant $\mu < \infty$ such that the following holds. Let W be a binary-input memoryless output-symmetric channel with capacity $I(W)$. Then there exists $a_W < \infty$ such that for all $\varepsilon > 0$ and all powers of two $N \geqslant a_W (1/\varepsilon)^{\mu}$, there is a deterministic $\text{poly}(N)$ time construction of a binary linear code of block length N and*

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[1]One can avoid the brute force search for a good inner code by using a small ensemble of capacity-achieving codes in a Justesen-style construction [6]. But this will require taking the outer code length $n_0 > \exp(1/\varepsilon^2)$, causing a large delay.

[2]Spatially coupled LDPC codes were also recently shown to achieve the capacity of general BIS channels [7]. This construction gives a random code ensemble as opposed to a specific code, and as far as we know, rigorous bounds on the code length as a function of gap to capacity are not available.

rate at least $I(W) - \varepsilon$ and a deterministic $N \cdot \text{poly}(\log N)$ time decoding algorithm for the code with block error probability at most $2^{-N^{0.49}}$ for communication over W.

Remarks:

1) Using our results about polar codes, we can also construct codes of rate $I(W) - \varepsilon$ with $2^{-\Omega_\varepsilon(N)}$ block error probability (similar to Shannon's theorem) with similar claims about the construction and decoding complexity. The idea is to concatenate an outer code that can correct a small fraction of worst-case errors with a capacity-achieving polar code of dimension $\text{poly}(1/\varepsilon)$ as the inner code. A similar idea with outer Reed-Solomon codes yielding $2^{-\Omega(N/\text{poly}(\log N))}$ block error probability is described in [8].

2) The construction time in Theorem 1 can be made $\text{poly}(1/\varepsilon) + \mathcal{O}(N \log N)$. As our main focus is on the finite-length behavior when N is also $\text{poly}(1/\varepsilon)$, we are content with stating the $\text{poly}(N)$ claim above.

Showing that polar codes have a gap to capacity that is polynomially small in $1/N$ is our principal contribution. The decoding algorithm remains the same successive cancellation decoder of Arıkan [1]. The proof of efficient constructibility follows the approach, originally due to Tal and Vardy [9], of approximating the channels corresponding to different input bits seen at the decoder by a degraded version with a smaller output alphabet. The approximation error of this process and some of its variants were analyzed in [10]. We consider and analyze a somewhat simpler degrading process. One slight subtlety here is that we can only estimate the channel's Bhattacharyya parameter within error that is polynomial in $1/N$ in $\text{poly}(N)$ time, which will limit the analysis to an inverse polynomial block error probability. To get a block error probability of $2^{-N^{0.49}}$ we use a two step construction method that follows our analysis of the polarization process. As a bonus, this gives the better construction time alluded to in the second remark above.

Prior to our work, it was known that the block error probability of successive cancellation decoding of polar codes is bounded by $2^{-N^{0.49}}$ for rate approaching $I(W)$ in the limit of $N \to \infty$ [11]. However, the underlying analysis found in [11], which depended on the martingale convergence theorems, did not offer any bounds on the finite-length convergence to capacity, i.e., the block length N required for the rate to be within ε of the capacity $I(W)$. To quote from the introduction of the recent breakthrough on spatially coupled LDPC codes [7]:

"There are perhaps only two areas in which polar codes could be further improved. First, for polar codes the convergence of their performance to the asymptotic limit is slow. Currently no rigorous statements regarding this convergence for

the general case are known. But "calculations" suggest that, for a fixed desired error probability, the required block length scales like $1/\delta^{\mu}$, where δ is the additive gap to capacity and where μ depends on the channel and has a value around 4."[3]

The above-mentioned heuristic calculations are based on "scaling laws" and presented in [12]. We will return to the topic of scaling laws in Section I-D on related work.

We note that upper bounds on the block length N as a function of gap ε to capacity are crucial, as without those we cannot estimate the complexity of communicating at rates within ε of capacity. Knowing that the asymptotic complexity is $\mathcal{O}(N \log N)$ for large N by itself is insufficient (for example, to claim that polar codes are better than concatenated codes) as we do not know how large N has to be! While an explicit value of μ in Theorem 1 can be calculated, it will be rather large, and obtaining better bounds on μ, perhaps closer to the empirically suggested bound of ≈ 4, is an interesting open problem[4].

C. Techniques

Let us first briefly discuss the concept of polarization in Arıkan's work, and then turn to aspects of our work. More formal background on Arıkan's construction of polar codes appears in Section III (with slightly different and notation that is more conventional in the polar coding literature). A good, easy to read, reference on polar codes is the recent survey by Şaşoğlu [14].

Fix W to be an arbitrary symmetric channel. If we have a capacity-achieving binary linear code C of block length N for W, then it is not hard to see that by padding the generator matrix of C one can get an $N \times N$ invertible matrix A_N with the following *polarization property*. Let $\mathbf{u} \in \{0,1\}^N$ be a uniformly random (column) vector. Given the output $\mathbf{y}$ of W when the N bits $\mathbf{x} = A_N \mathbf{u}$ are transmitted on it, for a $1 - o(1)$ fraction of bits u_i, its conditional entropy given $\mathbf{y}$ and the previous bits $u_1, \dots, u_{i-1}$ is either close to 0 (i.e., that bit can be determined with good probability) or close to 1 (i.e., that bit remains random). Since the conditional entropies of $\mathbf{u}$ given $\mathbf{y}$ and $\mathbf{x}$ given $\mathbf{y}$ are equal to each other, and the latter is $\approx (1 - I(W))N$, the fraction of bits u_i for which the conditional entropy given $\mathbf{y}$ and the previous bits $u_1, \dots, u_{i-1}$ is ≈ 0 (resp. ≈ 1) is $\approx I(W)$ (resp. $\approx 1 - I(W)$).

[3] The second aspect concerns *universality*: the design of polar codes depends on the channel being used, and the same code may not achieve capacity over a non-trivial class of channels.

[4] While we were completing the writeup of this paper and circulating a draft, we learned about a recent independently-derived result in [13] stating that $\mu \approx 6$ would suffice for block error probabilities bounded by an inverse polynomial. Our analysis primarily focuses on the $2^{-N^{.49}}$ block error probability result.

Arıkan gave a recursive construction of such a polarizing matrix A_N for $N = 2^n$: $A_N = G_2^{\otimes n} B_n$ where $G_2 = \left(\begin{smallmatrix} 1 & 1 \\ 0 & 1 \end{smallmatrix} \right)$ and B_n is a permutation matrix (for the bit-reversal permutation). In addition, he showed that the recursive structure of the matrix implied the existence of an efficiently decodable capacity-achieving code. The construction of this code amounts to figuring out which input bit positions have conditional entropy ≈ 0, and which don't (the message bits u_i corresponding to the latter positions are "frozen" to 0).

The proof that A_N has the above polarization property proceeds by working with the Bhattacharyya parameters $Z_n(i) \in [0,1]$ associated with decoding u_i from $\mathbf{y}$ and $u_1, \dots, u_{i-1}$. This quantity is the Hellinger affinity between the output distributions when $u_i = 0$ and $u_i = 1$. The values of the Bhattacharyya parameter of the 2^n bit positions at the n'th level can be viewed as a random variable Z_n (induced by the uniform distribution on the 2^n positions). The simple recursive construction of A_N enabled Arıkan to proved that the sequence of random variables $Z_0, Z_1, Z_2, \dots$ form a supermartingale. In particular, Z_{n+1} equals Z_n^2 with probability $1/2$ and is at most $2Z_n - Z_n^2$ with probability $1/2$. [5]

One can think the evolution of the Bhattacharyya parameter as a stochastic process on the infinite binary tree, where in each step we branch left or right with probability $1/2$. The polarization property is then established by invoking the martingale convergence theorem for supermartingales. The martingale convergence theorem implies that $\lim_{n \to \infty} |Z_{n+1} - Z_n| = 0$, which in this specific case also implies $\lim_{n \to \infty} Z_n(1 - Z_n) = 0$ or in other words polarization of Z_n to 0 or 1 for $n \to \infty$. However, it does *not* yield any effective bounds on the *speed* at which polarization occurs. In particular, it does not say how large n must be as a function of ε before $\mathbb{E}[Z_n(1 - Z_n)] \leqslant \varepsilon$; such a bound is necessary, though not sufficient, to get codes of block length 2^n with rate within ε of capacity.

In this work, we first work with the entropy of the channels $W_n^{(i)}$ associated with decoding the i'th bit, namely $H(W_n^{(i)}) = H(u_i \mid \mathbf{y}, u_1, u_2, \dots, u_{i-1})$ to prove that these values polarize to 0 and 1 exponentially fast in the number of steps n. Formally, we prove that for $n = O(\log 1/\varepsilon)$, $H_n \in (\varepsilon, 1 - \varepsilon)$ with probability at most ε, where H_n is the random variable associated with the entropy values $H(W_n^{(i)})$ at the n'th level. As the Bhattacharyya parameter is within a square root factor of the entropy, we get a similar claim about Z_n. The advantage in working with the entropy instead of the Bhattacharyya parameter is that the entropy forms a martingale, so that given H_n, the two possible values of H_{n+1} are $H_n \pm \alpha$ for some $\alpha \geqslant 0$. We show

[5] For the special case of the binary erasure channel, the Bhattacharyya parameters simply equal the probability that the bit is unknown. In this case, the upper bound of $2Z_n - Z_n^2$ becomes an exact bound, and the Z_i's form a martingale.

that these two values are sufficiently separated, specifically that $\alpha \geqslant \frac{3}{4} H_n(1 - H_n)$. Thus, unless H_n is very close to 0 or 1, the two new values have a sizeable difference. We use this to show that $\mathbb{E}\big[\sqrt{H_n(1 - H_n)}\big]$ decreases by a constant factor in each step, which implies the desired exponential decay in $H_n(1 - H_n)$ and therefore also $Z_n(1 - Z_n)$. [6]

The above bound is itself, however, not enough to prove Theorem 1. What one needs is *fine polarization*, where the smallest $\approx I(W)N$ values among $Z_n(i)$ all *add up* to a quantity that tends to 0 for large N (in fact, this sum should be at most $2^{-N^{0.49}}$ if we want the block error probability claimed in Theorem 1). To establish this, we use that in further steps, Z_{n+1} reduces rapidly to Z_n^2 with probability $1/2$, together with Chernoff-bound arguments (similar to [11]) to bootstrap the rough polarization of the first step to a fine polarization that suffices to bound the block error decodability.

Our analysis is elementary and self-contained, and does not use the martingale convergence theorem. The ingredients in our analysis were all present explicitly or implicitly in various previous works. However, it appears that their combination to imply a polynomial convergence to capacity has not been observed before, as evidenced by the explicit mention of this as an open problem in the literature, eg. [15, Section 6.6], [7, Section Ia], [16, Section 1.3], and [9, Section I] (see the discussion following Corollary 2).

D. Related work

The simplicity and elegance of the construction of polar codes, and their wide applicability to a range of classic information theory problems, have made them a popular choice in the recent literature. Here we only briefly discuss aspects close to our focus on the speed of polarization.

Starting with Arıkan's original paper, the "rate of polarization" has been studied in several works. However, this refers to something different than our focus; this is why we deliberately use the term "speed of polarization" to refer to the question of how large n needs to be before, say, Z_n is in the range $(\varepsilon, 1 - \varepsilon)$ with probability ε. The rate of polarization refers to pinpointing a function Υ with $\Upsilon(n) \to 0$ for large n such that $\lim_{n \to \infty} \Pr[Z_n \leqslant \Upsilon(n)] = I(W)$. Arıkan proved that one can take $\Upsilon(n) = O(2^{-5n/4})$ [1], and later Arıkan and Telatar established that one can take $\Upsilon(n) = 2^{-2^{\beta n}}$ for any $\beta < 1/2$ [11]. Further they proved that for $\gamma > 1/2$, $\lim_{n \to \infty} \Pr[Z_n \leqslant 2^{-2^{\gamma n}}] = 0$. This determined the rate at which the Bhattacharyya parameters of the "noiseless" channels polarize to 0 in the limit of larger n. More fine grained bounds on this asymptotic rate of polarization as a function of the code rate were obtained in [17].

For our purpose, to get a finite-length statement about the performance of polar codes, we need to understand the speed at which $\Pr[Z_n \leqslant \Upsilon(n)]$ approaches the limit $I(W)$ as n grows (any function Υ with $\Upsilon(n) = o(1/2^n)$ will do, though we get the right $2^{-2^{0.49n}}$ type decay).

Restated in our terminology, in [18] the authors prove the following "negative result" concerning gap to capacity: for polar coding with successive cancellation (SC) decoding to have vanishing decoding error probability at rates within ε of capacity, the block length has to be *at least* $(1/\varepsilon)^{3.553}$. (A slight caveat is that this uses the sum of the error probabilities of the well-polarized channels as a proxy for the block error probability, whereas in fact this sum is only an upper bound on the decoding error probability of the SC decoder.)

Also related to the gap to capacity question is the work on "scaling laws," which is inspired by the behavior of systems undergoing a phase transition in statistical physics. In coding theory, scaling laws were suggested and studied in the context of iterative decoding of LDPC codes in [19]. In that context, for a channel with capacity C, the scaling law posits the existence of an exponent μ such that the block error probability $P_e(N, R)$ as a function of block length N and rate R tends in the limit of $N \to \infty$ while fixing $N^{1/\mu}(C - R) = x$, to $f(x)$ where f is some function that decreases smoothly from 1 to 0 as its argument changes from $-\infty$ to $+\infty$. Coming back to polar codes, in [12], the authors make a *Scaling Assumption* that the probability $Q_n(x)$ that Z_n exceeds x is such that $\lim_{n \to \infty} N^{1/\mu} Q_n(x)$ exists and equals a function $Q(x)$. Under this assumption, they use simulations to numerically estimate $\mu \approx 3.627$ for the BEC. Using the small x asymptotics of $Q(x)$ suggested by the numerical data, they predict an $\approx (1/\varepsilon)^\mu$ upper bound on the block length as a function of the gap ε to capacity for the BEC. For general channels, under the heuristic assumption that the densities of log-likelihood ratios behave like Gaussians, an exponent of $\mu \approx 4.001$ is suggested for the Scaling Assumption. However, to the best of our knowledge, it does not appear that one can get a rigorous upper bound on block length N as a function of the gap to capacity via these methods.

II. PRELIMINARIES

We will work over a binary input alphabet $\mathcal{B} = \{0, 1\}$. Let $W : \mathcal{B} \to \mathcal{Y}$ be a binary-input memoryless symmetric channel with finite output alphabet $\mathcal{Y}$ and transition probabilities $\{W(y|x) : x \in \mathcal{B}, y \in \mathcal{Y}\}$. A binary-input channel is symmetric when the two rows of the transition probability matrix are permutations of each other; i.e., there exists a bijective mapping $\sigma : \mathcal{Y} \mapsto \mathcal{Y}$ where $\sigma = \sigma^{-1}$ and $W(y|0) = W(\sigma(y)|1)$ for all y. Both the binary erasure and binary symmetric channels are examples of symmetric channels.

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[6]We note that one can also prove directly that $\mathbb{E}[Z_n(1 - Z_n)]$ decreases by a constant factor in each step and an earlier version of this paper (and independently [13]) used this approach. The analysis presented here in terms of H_n is cleaner and more intuitive in our opinion.

Let X represent a uniformly distributed binary random variable, and let Y represent the result of sending X through the channel W.

The entropy of the channel W, denote $H(W)$, is defined as the entropy of X, the input, given the output Y, i.e., $H(W) = H(X|Y)$. It represents how much uncertainty there is in the input of the channel given the output of the channel. The mutual information of W, sometimes known as the capacity, and denoted $I(W)$, is defined as the mutual information between X and Y when the input distribution X is uniform:

$$I(W) = I(X;Y)$$
$$= 1 - H(X|Y)$$
$$= 1 - H(W) \ .$$

We have $0 \leqslant I(W) \leqslant 1$, with a larger value meaning a less noisy channel. As the mutual information expression is difficult to work with directly, we will often refer to the Bhattacharyya parameter of W as a proxy for the quality of the channel:

$$Z(W) = \sum_{y \in \mathcal{Y}} \sqrt{W(y|0)W(y|1)} \ .$$

This quantity is a natural one to capture the similarity between the channel outputs when the input is 0 and 1: $Z(W)$ is simply the dot product between the unit vectors obtained by taking the square root of the output distributions under input 0 and 1 (which is also called the Hellinger affinity between these distributions).

Intuitively, the Bhattacharyya parameter $Z(W)$ should be near 0 when $H(W)$ is near 0 (meaning that it is easy to ascertain the input of a channel given the output), and conversely, $Z(W)$ is near 1 when $H(W)$ is near 1. This intuition is quantified by the following expression (where the upper bound is from [15, Lemma 1.5] and the lower bound is from [20]):

$$Z(W)^2 \leqslant H(W) \leqslant Z(W) \ . \tag{1}$$

Given a single output $y \in \mathcal{Y}$ from a channel W, we would like to be able to map it back to X, the input to the channel. The most obvious way to do this is by using the maximum-likelihood decoder:

$$\hat{X} = \operatorname*{argmax}_{x \in \mathcal{B}} \Pr(x|y) = \operatorname*{argmax}_{x \in \mathcal{B}} W(y|x)$$

where a decoding error is declared if there is a tie. Thus, the probability of error for a uniform input bit under maximum likelihood decoding is

$$P_e(W) = \Pr(\hat{X} \neq X)$$
$$= \frac{1}{2} \sum_{x \in \mathcal{B}} \sum_{y \in \mathcal{Y}} W(y|x) \, \mathbf{1}_{W(y|x) \leqslant W(y|x \oplus 1)}$$

where $\mathbf{1}_x$ denotes the indicator function of x. Directly from this expression, we can conclude

$$P_e(W) \leqslant Z(W) \tag{2}$$

since $\mathbf{1}_{W(y|x) \leqslant W(y|x \oplus 1)} \leqslant \sqrt{W(y|x \oplus 1)}/\sqrt{W(y|x)}$, and the channel is symmetric (so the sum over $x \in \mathcal{B}$ and the $1/2$ cancel out). Thus, the Bhattacharyya parameter $Z(W)$ also bounds the error probability of maximum likelihood decoding based on a single use of the channel W.

III. Polar codes

A. Construction preliminaries

This is a short primer on the motivations and techniques behind polar coding, following [1], [14]. Consider a family of invertible linear transformations $G_n : \mathcal{B}^{2^n} \to \mathcal{B}^{2^n}$ defined recursively as follows: $G_0 = [1]$ and for a $2N$-bit vector $u = (u_0, u_1, \ldots, u_{2N-1})$ with $N = 2^n$, we define

$$G_{n+1}u = G_n(u_0 \oplus u_1, u_2 \oplus u_3, \ldots, u_{2N-2} \oplus u_{2N-1}) \tag{3}$$
$$\circ \ G_n(u_1, u_3, u_5, \ldots, u_{2N-1})$$

where $\circ$ is the vector concatenation operator. More explicitly, this construction can be shown to be equivalent to the explicit form $G_n = K^{\otimes n} B_n$ (see [1, Sec. VII]) where B_n is the $2^n \times 2^n$ bit-reversal permutation matrix for n-bit strings, $K = \begin{bmatrix} 1 & 1 \\ 0 & 1 \end{bmatrix}$ and $\otimes$ denotes the Kronecker product.

Suppose we use the matrix G_n to encode a $N = 2^n$-size vector U, $X = G_n U$, and then transmit X over a binary symmetric channel W. It can be shown with a Martingale Convergence Theorem-based proof [1] that for all $\varepsilon > 0$,

$$\lim_{N \to \infty} \Pr_i \left[H(U_i|U_0^{i-1}, Y_0^{N-1}) < \varepsilon \right] = I(W). \tag{4}$$

where the notation U_i^j denotes the subvector $(U_i, U_{i+1}, \ldots, U_j)$.

In words, there exists a *good set* of indices i so that for all elements in this set, given all of the outputs from the channel and (correct) decodings of all of the bits indexed less than i, the value of U_i can be ascertained with low probability of error (as it is a low-entropy random variable).

For every element that is outside of the good set, we do not have this guarantee; this suggests a encoding technique wherein we "freeze" all indices outside of this good set to a certain predefined value (0 will do). We call the indices that are not in the good set as the *frozen* set.

B. Successive cancellation decoder

The above distinction between good indices and frozen indices suggests a successive cancellation decoding technique where if the index is in the good set, we output the maximum-likelihood bit (which has low probability of being wrong due to the low entropy) or if the index is in the

frozen set, we output the predetermined bit (which has zero probability of being incorrect). A sketch of such a successive cancellation decoder is presented in Algorithm 1.

Definition 1. *A polar code with frozen set* $F \subset \{0, 1, \dots, N-1\}$ *is defined as*

$$C_F = \{G_n u \mid u \in \{0,1\}^N, u_F = 0\} \ .$$

Algorithm 1: Successive cancellation decoder

input : y_0^{N-1}, F, W
output: u_0^{K-1}
1 $\hat{u} \leftarrow$ zero vector of size N
2 **for** $i \in 0..N-1$ **do**
3   **if** $i \in F$ **then**
4        $\hat{u}_i \leftarrow 0$
5   **else**
6   **if** $\frac{\Pr(U_i=0|U_0^{i-1}=\hat{u}_0^{i-1}, Y_0^{N-1}=y_0^{N-1})}{\Pr(U_i=1|U_0^{i-1}=\hat{u}_0^{i-1}, Y_0^{N-1}=y_0^{N-1})} > 1$ **then**
7           $\hat{u}_i \leftarrow 0$
8   **else**
9           $\hat{u}_i \leftarrow 1$

10 **return** $\hat{u}_{\overline{F}}$

Remark. The runtime of the algorithm can be improved to $\mathcal{O}(N \log N)$ by computing the probabilities on line 6 with a divide-and-conquer approach as in [1]. We note that this runtime bound assumes constant-time arithmetic; consideration of n-bit arithmetic relaxes this bound to $\mathcal{O}(N\text{polylog}(N))$. For a treatment of more aggressive quantizations, see [13, Chapter 6].

By (4), if we take F to be the positions with conditional entropy exceeding ε, the rate of such a code would approach $I(W)$ in the limit $N \to \infty$.

To simplify the probability calculation (as seen on line 6 of Algorithm 1 and explained further in the comments), it is useful to consider the induced channel seen by each bit, $W_n^{(i)} : \mathcal{B} \to \mathcal{Y}^N \times \mathcal{B}^i$, for $0 \leqslant i \leqslant 2^n - 1$. Here, we are trying to ascertain the most probable value of the input bit U_i by considering the output from all channels Y_0^{N-1} and the (decoded) input from all channels before index i. Since the probability of decoding error at every step is bounded above by the corresponding Bhattacharyya parameter Z by (2), we can examine $Z(W_n^{(i)})$ as a proxy for $P_e(W_n^{(i)})$.

It will be useful to redefine $W_n^{(i)}$ recursively both to bound the evolution of $Z(W_n^{(i)})$ and to facilitate the computation. Consider the two transformations $^-$ and $^+$ defined as follows:

$$W^-(y_1, y_2|x_1) = \sum_{x_2 \in \mathcal{B}} \frac{1}{2} W(y_1|x_1 \oplus x_2) W(y_2|x_2) \quad (5)$$

and

$$W^+(y_1, y_2, x_1|x_2) = \frac{1}{2} W(y_1|x_1 \oplus x_2) W(y_2|x_2). \quad (6)$$

This process (5) and (6) preserves information in the sense that

$$I(W^-) + I(W^+) = 2I(W), \quad (7)$$

which follows by the chain rule of mutual information, as (suppose X_1 is the input seen at W^- and X_2 is the input seen at W^+ and Y_1, Y_2 are the corresponding output variables)

$$I(W^-) + I(W^+) = I(X_1; Y_1, Y_2) + I(X_2; Y_1, Y_2|X_1)$$
$$= I(X_1, X_2; Y_1, Y_2) = 2I(W).$$

We also associate $^-$ with a "downgrading" transformation and $^+$ with an "upgrading" transformation, as $I(W^-) \leqslant I(W) \leqslant I(W^+)$.

Tying the operations $^-$ and $^+$ back to $Z(W_n^{(i)})$, we notice that $W^- = W_1^{(0)}$ (the transformation $^-$ adds uniformly distributed noise from another input x_2, which is equivalent to the induced channel seen by the 0th bit) and $W^+ = W_1^{(1)}$ (where here we clearly have the other input bit). More generally, by the recursive construction (3), one can conclude that the $W_n^{(i)}$ process can be redefined in a recursive manner as

$$W_{n+1}^{(i)} = \begin{cases} \left(W_n^{(\lfloor i/2 \rfloor)}\right)^- & \text{if } i \text{ is even} \\ \left(W_n^{(\lfloor i/2 \rfloor)}\right)^+ & \text{if } i \text{ is odd} \end{cases} \quad (8)$$

with the base channel $W_0^{(0)} = W$.

The evolution of $I(W^+)$ and $I(W^-)$ is difficult to analyze, but we will see in the next section that we can adequately bound $Z(W^+)$ and $Z(W^-)$ as a proxy. Such bounds are sufficient for analyzing our decoder, as we can bound the block error probability obtained by the successive cancellation decoder described in algorithm 1 with bounds on the Bhattacharyya parameters of the subchannels. The probability of the ith (not frozen) bit being misdecoded by the algorithm, given the channel outputs and the input bits with index less than i, is bounded above by $Z(W_n^{(i)})$ by equation (2). This observation, with the union bound, immediately gives the following lemma.

Lemma 2. *The block error probability of Algorithm 1 on a polar code C of length n with frozen set F is bounded above by the sum of the Bhattacharyya parameters $\sum_{i \in \overline{F}} Z(W_n^{(i)})$.*

C. Bounds on $Z(W^-)$ and $Z(W^+)$

The proof of these well-known bounds can be found in [1], [15].

Proposition 3. $Z(W^+) = Z(W)^2$ *for all binary symmetric channels W.*

Proposition 4. $Z(W^-) \leqslant 2Z(W) - Z(W)^2$ *for all binary symmetric channels W, with equality if the channel W is an erasure channel.*

IV. Speed of Polarization

Our first goal is to show that for some $m = \mathcal{O}(\log(1/\varepsilon))$, we have that $\Pr_i[Z(W_m^{(i)}) \leqslant 2^{-\mathcal{O}(m)}] \geqslant I(W) - \varepsilon$ (the channel is "roughly" polarized). We will then use this rough polarization result to show that, for some $n = \mathcal{O}(\log(1/\varepsilon))$, "fine" polarization occurs: $\Pr_i[Z(W_n^{(i)}) \leqslant 2^{-2^{\beta n}}] \geqslant I(W) - \varepsilon$. This approach is similar to the bootstrapping method used in [21].

A. Rough polarization

We give a formal statement of rough polarization in the proposition below. A similar statement can be constructed for binary erasure channels (as opposed to general symmetric channels) with a much simpler proof; this analysis can be found in the full version of the paper.

Proposition 5. *There is an absolute constant $\Lambda < 1$ such that the following holds. For all $\rho \in (\Lambda, 1)$, there exists a constant c_ρ such that for all binary-input symmetric channels W, all $\varepsilon > 0$ and $m \geqslant b_\rho \log(1/\varepsilon)$, there exists a* roughly polarized *set*

$$\mathcal{W}_r \subset \mathcal{W} \triangleq \{W_m^{(i)} : 0 \leqslant i \leqslant 2^m - 1\} \quad (9)$$

such that for all $M \in \mathcal{W}_r$, $Z(M) \leqslant 2\rho^m$ and $\Pr_i(W_m^{(i)} \in \mathcal{W}_r) \geqslant I(W) - \varepsilon$.

We first offer the following quantitative bound on the evolution of each step of the polarization process.

Lemma 6. *For all channels W, we have $H(W^+) \leqslant H(W) - \alpha(W)$ and $H(W^-) \geqslant H(W) + \alpha(W)$ for $\alpha(W) = \theta H(W)(1 - H(W))$, where θ is a constant greater than $3/4$.*

Proof: Let

$$\hat{\theta} = \inf_W \frac{H(W^-) - H(W)}{H(W)(1 - H(W))},$$

where the minimization is done over all binary-input symmetric channels W. Expanding the definition of the $^-$ transform, obtain

$$H(W^-) - H(W) = H(X_1 + X_2 | Y_1, Y_2) - H(X_1 | Y_1) \quad (10)$$

where X_1, X_2 are uniformly distributed random bits, Y_1 and Y_2 in the first expression are distributed according to the transition probabilities of W^- and Y_1 is distributed according to the transition probabilities of W.

[14, Lemma 2.2] implies that if (X_1, Y_1) and (X_2, Y_2) are independent pairs of discrete random variables with $X_1, X_2 \in \mathcal{B}$ and $H(X_1 | Y_1) = H(X_2 | Y_2) = \alpha$, we have

$$H(X_1 + X_2 | Y_1, Y_2) - H(X_1 | Y_1) \geqslant \varepsilon(\alpha),$$

where $\varepsilon(\alpha) = h(2h^{-1}(\alpha)(1 - h^{-1}(\alpha))) - \alpha$ (here, h is the binary entropy function and h^{-1} is its inverse). Substituting the above in (10), we can write

$$H(W^-) - H(W) \geqslant h(2h^{-1}(\alpha)(1 - h^{-1}(\alpha))) - \alpha. \quad (11)$$

We can therefore bound the desired expression by numerically minimizing the expression $\frac{h(2x(1-x)) - h(x)}{h(x)(1 - h(x))}$ over $x \in (0, 1/2)$ (the range of h^{-1}), which offers us $\hat{\theta} > .799$.

Since mutual information is conserved in our transformation (as stated in Equation (7)), we can conclude the lemma, as any $\theta < \hat{\theta}$ suffices for the statement to be true. ∎

We define the symmetric entropy of a channel as

$$T(W) = H(W)(1 - H(W)).$$

To relate $T(W_n^{(i)})$ back to $H(W_n^{(i)})$, it is useful to define the sets (where $\rho \in (0, 1)$): $A_\rho^g = \{i : H(W_n^{(i)}) \leqslant (1 - \sqrt{1 - 4\rho^n})/2\}$, $A_\rho^b = \{i : H(W_n^{(i)}) \geqslant (1 + \sqrt{1 - 4\rho^n})/2\}$, and $A_\rho = A_\rho^g \cup A_\rho^b$. We associate A_ρ^g with the "good" set (the set of i such that the entropy, and therefore probability of misdecoding, is small) and A_ρ^b with the "bad" set. We record the following useful approximations, both of which follow from $\sqrt{1 - 4\rho^n} \geqslant 1 - 4\rho^n$.

Fact 7. *For $i \in A_\rho^g$, $H(W_n^{(i)}) \leqslant 2\rho^n$, and for $i \in A_\rho^b$, $H(W_n^{(i)}) \geqslant 1 - 2\rho^n$.*

We first state a bound on the evolution of $\sqrt{T(W_{n+1}^{(i)})}$.

Lemma 8. *There exists a universal constant $\Lambda < 1$ such that*

$$\mathop{\mathbb{E}}_{i \bmod 2} \sqrt{T(W_{n+1}^{(i)})} \leqslant \Lambda \sqrt{T(W_n^{(\lfloor i/2 \rfloor)})},$$

where the meaning of the expectation is that we fix $\lfloor i/2 \rfloor$ and allow $i \bmod 2$ to vary.

Proof: Defining $h = H(W_n^{(i)})$, we have

$$\mathop{\mathbb{E}}_{i \bmod 2} \sqrt{T(W_{n+1}^{(i)})} = \frac{1}{2}\left(\sqrt{h(1 - h) + (1 - 2h)\alpha - \alpha^2}\right.$$
$$\left. + \sqrt{h(1 - h) - (1 - 2h)\alpha - \alpha^2}\right)$$
$$(12)$$

where $\alpha = H((W_n^{(i)})^-) - H(W_n^{(i)}) = H(W_n^{(i)}) - H((W_n^{(i)})^+)$. By symmetry, we can assume $h \leqslant 1/2$ without loss of generality, and we also know that $\alpha \geqslant \theta h(1 - h)$ from Lemma 6. We can write (where the expectation is taken

over $i \bmod 2$)

$$2\,\mathbb{E}\,\sqrt{T(W_{n+1}^{(i)})} \leqslant \sqrt{h(1-h)+(1-2h)\alpha}$$
$$+ \sqrt{h(1-h)-(1-2h)\alpha}$$
$$\leqslant \sqrt{h(1-h)} - \frac{((1-2h)\alpha)^2}{4\,((h(1-h))^{3/2}}$$
$$\leqslant \sqrt{h(1-h)} - \frac{\theta^2}{4}(1-2h)\sqrt{h(1-h)}$$

where the second line is a Taylor expansion around $h(1-h)$; the coefficients on the odd order terms are 0 and the coefficients on the even order terms are negative, so we can truncate the series and maintain the inequality. This analysis gives the desired result for whenever $1-2h$ is greater than an absolute constant. For clarity of analysis, let us fix a concrete constant $1-2h \geqslant 1/100$.

We can therefore focus on the case where $1-2h < 1/100$, which implies $h \in [99/200, 1/2]$. Continuing, we have

$$\alpha \geqslant \theta h(1-h) \geqslant 99\theta/400, \tag{13}$$

an absolute constant bounded away from zero. We can also write $\alpha \geqslant 2(1-2h)$ as $2(1-2h) \leqslant 1/50$, which is less than $99\theta/400$, since $\theta > 3/4$ from Lemma 6. This expression implies

$$(1-2h)\alpha - \alpha^2 \leqslant -\alpha^2/2$$

which, when inserted into (12), offers

$$\mathbb{E}_{i \bmod 2}\sqrt{T(W_{n+1}^{(i)})} \leqslant \sqrt{h(1-h)-\frac{\alpha^2}{2}}.$$

This implies the existence of a $\Lambda < 1$, since α is bounded away from zero in (13), $h(1-h) \leqslant \frac{1}{4}$, and the function $\frac{\sqrt{x-c}}{\sqrt{x}}$ is increasing for positive c and $x > c$. ∎

Corollary 9. *Taking Λ as defined in Lemma 8,* $\Pr_i[T(W_n^{(i)}) \geqslant \alpha^n] \leqslant \frac{1}{2}\left(\frac{\Lambda^2}{\alpha}\right)^{n/2}$

Proof: Clearly we have

$$\mathbb{E}_i\sqrt{T(W_{n+1}^{(i)})} \leqslant \Lambda^n\sqrt{T(W)} \leqslant \Lambda^n \cdot \frac{1}{2}$$

and we can therefore use Markov's inequality to obtain the desired consequence. ∎

We are now in a position where we can conclude Proposition 5.

Proof of Proposition 5: We have

$$\Pr(\overline{A_\rho})\max_{i \in \overline{A_\rho}}(I(W_n^{(i)})) + \Pr(A_\rho^b)\max_{i \in A_\rho^b}I(W_n^{(i)}) +$$
$$\Pr(A_\rho^g)\max_{i \in A_\rho^g}I(W_n^{(i)}) \geqslant \mathbb{E}_i(I(W_n^{(i)})) = I(W) \tag{14}$$

where the last equality follows by the conservation of mutual information in our transformation as stated in equation (7).

As $\min_{i \in A_\rho^b}H(W_n^{(i)}) \geqslant 1 - 2\rho^n$ by Fact 7, we have $\max_{i \in A_\rho^b}I(W_n^{(i)}) \leqslant 2\rho^n$. Using this together with equation (14), obtain

$$\Pr(\overline{A_\rho}) + \Pr(A_\rho^b) \cdot 2\rho^n + \Pr(A_\rho^g) \geqslant I(W)$$

where we used the trivial inequality (for binary-input channels) $I(W_n^{(i)}) \leqslant 1$ for every i. Rearranging terms, using the bounds $\Pr(\overline{A_\rho}) \leqslant \frac{1}{2}(\Lambda^2/\rho)^{n/2}$ from Corollary 9 and $H(W_n^{(i)}) \leqslant 2\rho^n$ for $i \in A_\rho^g$ from Fact 7, we get

$$\Pr_i[H(W_m^{(i)}) \leqslant 2\rho^m] \geqslant \Pr(A_\rho^g)$$
$$\geqslant I(W) - \frac{1}{2}(\Lambda^2/\rho)^{m/2} - 2\rho^m . \tag{15}$$

Clearly, if $\rho > \Lambda^2$, there is a constant b_ρ such that $m \geqslant b_\rho \log(1/\varepsilon)$ implies that the above lower bound is at least $I(W) - \varepsilon$. We conclude our analysis by noting that $Z(W) \leqslant \sqrt{H(W)}$, as observed in (1), so that $\sqrt{\rho} > \Lambda$ can play the role of ρ for a lower bound similar to (15) on $\Pr_i[Z(W_m^{(i)}) \leqslant 2\kappa^m]$ for $\kappa \in (\Lambda, 1)$. ∎

B. Fine polarization

The following proposition formalizes what we mean by "fine polarization."

Proposition 10. *Given $\delta \in (0, 1/2)$, there exists a constant c_δ for all binary input memoryless channels W and $\varepsilon > 0$ such that if $n_0 > c_\delta \log(1/\varepsilon)$ then*

$$\Pr_i\left[Z(W_{n_0}^{(i)}) \leqslant 2^{-2^{\delta n_0}}\right] \geqslant I(W) - \varepsilon.$$

The proof of the proposition will require the following lemma to specify one of the constants.

Lemma 11. *For all $\gamma > 0$, $\beta \in (0, 1/2)$ and $\rho \in (0, 1)$, there exists a constant $\theta(\beta, \gamma, \rho)$ such that for all $\varepsilon \in (0, 1)$, if $m > \theta(\beta, \gamma, \rho) \cdot \log(2/\varepsilon)$, then*

$$\left(\frac{\lg(2/\rho)\gamma}{2}+1\right)\exp\left(-\frac{(1-2\beta)^2\lg(2/\rho)m}{2}\right) < \varepsilon/2 .$$

Proof: We can rewrite this expression as $c_1\exp(-c_2m) < \varepsilon$ for constants c_1, c_2 that are independent of ε and the result is clear. ∎

The following corollary will be useful in the next section, where we will deal with an approximation to the Bhattacharyya parameter. It relaxes the conditions on the polarized set from Proposition 5.

Corollary 12. *Proposition 10 still holds with a modified roughly polarized set (recall the definition of the roughly polarized set $\mathcal{W}_r$ from equation (9)) $\widetilde{\mathcal{W}_r}$ where $\widetilde{\mathcal{W}_r} \supset \mathcal{W}_r$ and $Z\left(\widetilde{\mathcal{W}_r}\right) \leqslant \sqrt{3\rho^m}$ (instead of $2\rho^m$) with a modified constant $\widetilde{c_\delta}$.*

The proofs of Proposition 10 and Corollary 12 can be found in the full version of the paper. The analysis is based on the argument in [21] and follows from a Chernoff-Hoeffding bound [22] on a modified process where Z_n either squares or doubles in every step.

V. Efficient Construction of Polar Codes

The construction of a polar code reduces to determining the frozen set of indices (the generator matrix then consists of columns of $G_n = K^{\otimes n} B_n$ indexed by the non-frozen positions). The core component of the efficient construction of a frozen set is estimating the Bhattacharyya parameters of the subchannels $W_n^{(i)}$. In the erasure case, this is simple because the evolution equation offered by Proposition 4 is exact. In the general case, the naïve calculation takes too much time: $W_n^{(i)}$ has an exponentially large output alphabet size in terms of $N = 2^n$.

Our goal, therefore, is to limit the alphabet size of $W_n^{(i)}$ while roughly maintaining the same Bhattacharyya parameter. With this sort of approach, we can select channels with relatively good Bhattacharyya parameters. The idea of approximating the channel behavior by degrading it via output symbol merging is due to [9] and variants of it were analyzed in [10]. The approach is also discussed in the survey [14, Section 3.3]. Since we can only achieve an inverse polynomial error in estimating the Bhattacharyya parameters with a polynomial alphabet, we use the estimation only up to the rough polarization step, and then use the explicit description of the subsequent good channels that is implicit in the proof of Proposition 10.

We note that revised versions of the Tal-Vardy work [9] also include a polynomial time algorithm for code construction by combining their methods with the analysis of [10]. However, as finite-length bounds on the speed of polarization were not available to them, they could not claim $\text{poly}(N/\varepsilon)$ construction time, but only $c_\varepsilon N$ time for some unspecified c_ε.

We will first state our binning algorithm, along with its properties, and then conclude the main theorem.

A. Binning Algorithm

For our binning, we deal with the marginal distributions of the input bit given an output symbol. A binary-input symmetric channel W defines a marginal probability distribution $W(y|x)$. We invert this conditioning to form the expression

$$p(0|y) = \Pr_x(x = 0 | W(x) = y) = \frac{1}{2} \frac{W(y|0)}{\Pr_x(W(x) = y)}$$

for a uniformly distributed input bit x. In addition, we introduce the one-argument form $p(y) = \Pr_x(W(x) = y)$ for the simple probability that the output is y given an uniformly distributed input bit x.

Algorithm 2: Binning algorithm

input : $W : \mathcal{B} \to \mathcal{Y}$, $k > 0$
output: $\widetilde{W} : \mathcal{B} \to \widetilde{\mathcal{Y}}$

1 Initialize new channel $\widetilde{W}$ with symbols $\tilde{y}_0, \tilde{y}_1 \ldots \tilde{y}_k$ with $\widetilde{W}(\tilde{y}|x) = 0$ for all $\tilde{y}$ and $x \in \mathcal{B}$

2 **for** $y \in \mathcal{Y}$ **do**

3    $p(0|y) \leftarrow \frac{1}{2} \frac{W(y|0)}{\Pr_x(W(x) = y)}$

4    $\widetilde{W}(\tilde{y}_{\lfloor kp(0|y) \rfloor}|0) \leftarrow \widetilde{W}(\tilde{y}_{\lfloor kp(0|y) \rfloor}|0) + W(y|0)$

5    $\widetilde{W}(\tilde{y}_{\lfloor kp(0|y) \rfloor}|1) \leftarrow \widetilde{W}(\tilde{y}_{\lfloor kp(0|y) \rfloor}|1) + W(y|1)$

6 **return** $\widetilde{W}$

Proposition 13. *For a binary-input symmetric channel $W : \mathcal{B} \to \mathcal{Y}$ and all $k > 0$, there exists a channel $\widetilde{W} : \mathcal{B} \to \widetilde{\mathcal{Y}}$ such that*

$$H(W) \leqslant H(\widetilde{W}) \leqslant H(W) + 2\lg(k)/k, \quad |\widetilde{\mathcal{Y}}| \leqslant k + 1 ,$$

and the channel transition probabilities, $\widetilde{W}(y|x)$, are computable, by Algorithm 2, in time polynomial in $|\mathcal{Y}|$ and k.

The proof of Proposition 13 is omitted here and can be found in the full version of the paper, as the details are somewhat mechanical. We note that a slightly different binning strategy [9] can achieve an approximation error of $O(1/k)$.

We will iteratively use the binning algorithm underlying Proposition 13 to select the best channels. The following corollary formalizes this.

Corollary 14. *Let $\widehat{W_n^{(i)}}$ indicate the result of using Algorithm 2 after every application of the evolution Equations (8); that is,*

$$\widehat{W_n^{(i)}} = \widetilde{\widetilde{\widetilde{W^{+^-}}}}^{\cdot^{\cdot^{\cdot}}}$$

where the $+$ or $-$ is chosen depending on the corresponding bit, starting from the least significant one, of the binary representation of $i \in \{0, 1, \ldots, 2^n - 1\}$. Then

$$H(W_n^{(i)}) \leqslant H\left(\widehat{W_n^{(i)}}\right) \leqslant H(W_n^{(i)}) + \frac{2^{n+2}\lg(k)}{k} .$$

The proof of Corollary 14 iterates the usage of Proposition 13 and can be found in the full version of the paper.

We are now in a position to prove our main theorem (Theorem 1).

Proof of Theorem 1: Fix an N that is a power of 2, and let $n_0 = \lg(N)$. Let $\mathcal{W}_r$ be the set of all channels $W_m^{(i)}$ such that $H(W_m^{(i)}) \leqslant 2\rho^m$, and $R(m)$ be the set of corresponding indices i. Utilizing the definition of $\,\widehat{}\,$ from Corollary 14 with $k = \left(\frac{2}{\rho}\right)^{2m}$, let $\widehat{\mathcal{W}_r}$ be the set of all channels $W_m^{(i)}$

such that $H\left(\widetilde{W_m^{(i)}}\right) \leqslant 3\rho^m$, and let $\hat{R}(m)$ be the set of corresponding indices i.

By Corollary 14 we can conclude that $i \in R(m)$ implies $i \in \hat{R}(m)$ because $Z(W_m^{(i)}) \leqslant 2\rho^m$ implies $H(W_m^{(i)}) \leqslant Z(W_n^{(i)}) \leqslant 2\rho^m$. This in turn implies $H(\widetilde{W_m^{(i)}}) \leqslant 3\rho^m$ by our choice of k and the approximation error guaranteed by Corollary 14. Therefore, we have

$$\Pr_{i<2^m}\left(i \in \hat{R}(m)\right) \geqslant \Pr_{i<2^m}\left(i \in R(m)\right)$$

and also that all $M \in \widehat{\mathcal{W}}_r$ satisfy $Z(M) \leqslant \sqrt{H(M)} \leqslant \sqrt{3\rho^m}$, where the former inequality is from (1).

The above analysis, combined with Corollary 12, implies that for some $m = \frac{1}{1+\gamma}n_0$ and some set $G(n_0 - m)$, there exists a set (where the notation $i_j^k = i/2^j \bmod 2^{k-j+1}$ means the integer with the binary representation of the jth through kth bits of i, inclusive):

$$\overline{\hat{F}_{n_0}} = \left\{ i \;\middle|\; \begin{array}{l} 0 \leqslant i \leqslant 2^{n_0} - 1, \\ i_0^{m-1} \in \hat{R}(m), \\ i_m^{n_0-1} \in G(n_0 - m) \end{array} \right\}$$

such that for $n_0 > c_\delta \log(1/\varepsilon)$, $\Pr_i(i \in \overline{\hat{F}_{n_0}}) \geqslant I(W) - \varepsilon$ and for all i in $\overline{\hat{F}_{n_0}}$, $Z(W_{n_0}^{(i)}) \leqslant 2^{-2^{\delta n_0}}$.

This implies that $\sum_{i \in \overline{\hat{F}_{n_0}}} Z(W_n^{(i)}) \leqslant N2^{-N^\delta}$. Taking $\delta = .499$ and $\mu = \tilde{c}_\delta$, we can conclude the existence of an a_W such that for $N \geqslant a_W(1/\varepsilon)^\mu$,

$$\sum_{i \in \overline{\hat{F}_{n_0}}} Z(W_n^{(i)}) \leqslant 2^{-N^{.49}},$$

as such μ satisfies the conditions of Corollary 12. We conclude by using Lemma 2: the block error probability of polar codes with a frozen set F under successive cancellation decoding is bounded by the sum of the Bhattacharyya parameters of the channels not in F.

We note that this set $\overline{\hat{F}_{n_0}}$ is computable in $\mathrm{poly}(1/\varepsilon, N)$ time: $\hat{R}(m)$ can be computed with poly-time Algorithm 2 with parameter $k \leqslant \mathrm{poly}(1/\varepsilon)$ and $G(n_0 - m)$ is computable in $\mathcal{O}(N)$ time: internally, the proof of Corollary 12 only requires the counting of 1 bits in various intervals to determine the "good" indices. ∎

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REFERENCES

[1] E. Arıkan, "Channel polarization: A method for constructing capacity-achieving codes for symmetric binary-input memoryless channels," *IEEE Transactions on Information Theory*, pp. 3051–3073, July 2009.

[2] J. Wolfowitz, "The coding of messages subject to chance errors," *Illinois J. Math.*, vol. 1, pp. 591–606, 1957.

[3] V. Strassen, "Asymptotische Abschatzungen in Shannon's Informationstheories," in *Trans. 3rd Prague Conf. Info. Theory*, 1962, pp. 689–723.

[4] G. D. Forney, "Concatenated codes," Ph.D. dissertation, Massachusetts Institute of Technology, 1967.

[5] V. Guruswami, "Iterative decoding of low-density parity check codes," *Bulletin of the EATCS*, vol. 90, pp. 53–88, 2006.

[6] J. Justesen, "A class of constructive asymptotically good algebraic codes," *IEEE Transactions on Information Theory*, vol. 18, no. 5, pp. 652–656, 1972.

[7] S. Kudekar, T. Richardson, and R. L. Urbanke, "Spatially coupled ensembles universally achieve capacity under belief propagation." in *ISIT*, 2012, pp. 453–457.

[8] M. Bakshi, S. Jaggi, and M. Effros, "Concatenated polar codes," in *ISIT*. IEEE, 2010, pp. 918–922.

[9] I. Tal and A. Vardy, "How to construct polar codes," *CoRR*, vol. abs/1105.6164v3, 2013.

[10] R. Pedarsani, S. H. Hassani, I. Tal, and E. Telatar, "On the construction of polar codes," in *ISIT*, 2011, pp. 11–15.

[11] E. Arıkan and E. Telatar, "On the rate of channel polarization," in *Proceedings of 2009 IEEE International Symposium on Information Theory*, 2009, pp. 1493–1495.

[12] S. B. Korada, A. Montanari, E. Telatar, and R. L. Urbanke, "An empirical scaling law for polar codes." in *ISIT*, 2010, pp. 884–888.

[13] S. H. Hassani, "Polarization and spatial coupling: Two techniques to boost performance," Ph.D. dissertation, École Polytechnique Fédérale De Lausanne, 2013.

[14] E. Şaşoğlu, "Polarization and polar codes," *Foundations and Trends in Communications and Information Theory*, vol. 8, no. 4, pp. 259–381, 2012.

[15] S. B. Korada, "Polar codes for channel and source coding," Ph.D. dissertation, École Polytechnique Fédérale De Lausanne, 2009.

[16] A. Shpilka, "Capacity achieving multiwrite WOM codes," *CoRR*, vol. abs/1209.1128, 2012.

[17] S. H. Hassani and R. L. Urbanke, "On the scaling of polar codes: I. the behavior of polarized channels." in *ISIT*, 2010, pp. 874–878.

[18] A. Goli, S. H. Hassani, and R. Urbanke, "Universal bounds on the scaling behavior of polar codes," in *Proceedings of 2012 IEEE International Symposium on Information Theory*, 2012, pp. 1957–1961.

[19] A. Amraoui, A. Montanari, T. J. Richardson, and R. L. Urbanke, "Finite-length scaling for iteratively decoded LDPC ensembles," *IEEE Transactions on Information Theory*, vol. 55, no. 2, pp. 473–498, 2009.

[20] E. Arıkan, "Source polarization," in *Proceedings of 2010 IEEE International Symposium on Information Theory*, 2010, pp. 899–903.

[21] E. Arıkan and E. Telatar, "On the rate of channel polarization," *CoRR*, vol. abs/0807.3806, 2008.

[22] W. Hoeffding, "Probability inequalities for sums of bounded random variables," *Journal of the American Statistical Association*, vol. 58, no. 301, pp. pp. 13–30, 1963. [Online]. Available: http://www.jstor.org/stable/2282952